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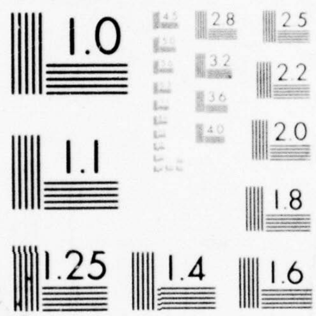

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MAY 1976

Varactor-Tuned Helical Resonators at UHF

UVE H. W. LAMMERS
MARTIN R. STIGLITZ

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ROME AIR DEVELOPMENT CENTER
AIR FORCE SYSTEMS COMMAND
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A JTIDS requirement calls for narrowing the front-end bandwidth of fast-tuning L-band receivers in order to improve their interference resistance. Filters suitable for this application have to be of high Q and compact design. Frequency tuning or switching within microseconds is required. The performance of optimally designed halfwave helical resonators is investigated with microwave varactors of Q > 4000 and Q > 10000 (at 50 MHz and 4V

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reverse bias) as tuning elements. Four filters were designed to cover sections of a frequency range 294 MHz to 1256 MHz. At the low end of this range the varactor Q is comparable to or better than that obtained from helical resonators of 1-in. shield diameter. At the high end of this range the varactor Q is such that there is a substantial degradation of overall resonator Q at maximum varactor capacitance (at 1V reverse bias). The compatibility of varactor-tuned helical filters with requirements of the JTIDS receivers is evaluated.

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1. INTRODUCTION

The Joint Tactical Information Distribution System (JTIDS) requires high-Q tunable filters for the uhf frequency range. The filters must be able to narrow the front-end bandwidth of fast-tuning receivers to match the instantaneous signal bandwidths as closely as possible. For a signal of bandwidth b that is switched within a frequency range larger than b, the resistance to interference is improved if the receiver of bandwidth b can follow the switching pattern and the interfering source cannot.

Stripline and helical resonators both provide low-cost compact filters for the uhf range but helical resonators have a higher unloaded quality factor Q_u. All hardware filters in the work reported here are halfwave helical structures tuned at the voltage maximum (center of the helix). The halfwave experimental design was selected because it allows a simple separation of input and output coupling circuits. If size is of paramount importance, a quarterwave resonator can be used.

To accomplish frequency changes within several microseconds while maintaining a high Q almost dictates the use of semiconductor tuning elements. Karp and Heynick[1] used multiple binary diode-switching of tuning capacitors. Their approach allows handling higher power but is more complex than the straight forward tuning with a capacitance diode that we investigated.

(Received for publication 18 May 1976)

1. Karp, A., and Heynick, L. N. (1975) UHF Electronically Tunable High-Power Filter, RADC-TR-75-220 [Final Tech. Rpt, September 1975].

Tuning varactors are limited both as to Q and maximum-to-minimum capacitance. The widest tuning range is obtained when the circuit connected to the varactor has a high L/C ratio. Although distributed circuits such as helical resonators have high Q, they have low L/C ratios; lumped inductors have lower stray capacitance and afford wider tuning but their Q does not approach that of helical resonators.

To analyze the varactor-tuned helical resonator, a lumped parallel-resonant circuit is used as a simple equivalent model. This model does not take into account any changes that varying capacitive loads cause in the field configuration along the distributed structure. The field configuration affects the input and output coupling circuits. We have to a certain extent experimentally investigated the effects of coupling changes by tuning the resonator with a low-loss variable capacitor. This has allowed us to at least qualitatively assess the effects of coupling changes that are superposed on the effects resulting from the nonideal characteristics of the varactor-tuning element.

2. SINGLE RESONANT CIRCUIT

Assume that the antenna represents a generator with impedance Z, and that the receiver front end represents a matched load Z to the antenna. The problem is to insert a passive filter between the antenna and the receiver; the insertion loss must be held to a minimum because of the given constant receiver gain and the reduction in signal-to-noise ratio.

A low insertion loss requires a tight coupling between generator, filter, and load, which is in direct contrast to the requirement for a narrow bandwidth. Tight coupling with the resonant circuit will lead to an intolerable reduction in circuit Q. The tradeoff between these two requirements and the inherent effect of nonideal varactor-loading is considered in Sections 2.2 and 2.3.

2.1 Parallel- Versus Series-resonant Filter

Either a parallel- or a series-resonant circuit can be inserted between the generator and the load. Figure 1 shows (a) a parallel-resonant circuit with inductance L_p, capacitance C_p, and resistance R_p; and (b) a series-resonant circuit with parameters L_s, C_s, and R_s. The following assumptions are made. Generator and load impedances Z are identical in both cases; both circuits have the same unloaded quality factor Q_u, the same resonance frequency ω, and the same loaded quality factor Q_L. By definition,

$$Q_L = \omega/[2(\omega - \omega_H)] , \qquad (1)$$

with ω_H the halfpower frequency below resonance. In the parallel-resonance case this frequency is characterized by

$$1/(\omega_H L_p) - \omega_H C_p = 2/Z + 1/R_p . \tag{2}$$

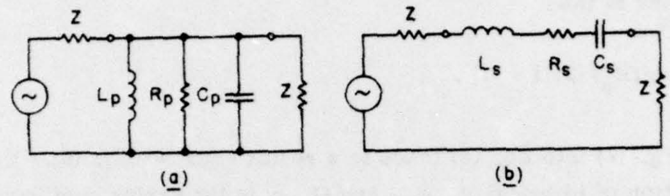

Figure 1. Parallel- and Series-resonant Filter Circuits

With the resonance condition $\omega^2 L_p C_p = 1$, and the unloaded $Q_u = R_p \omega C_p$, *it is* found that

$$K Z = 2 R_p , \tag{3}$$

where

$$K = \frac{\omega Q_u}{\omega_H} - \frac{\omega_H Q_u}{\omega} - 1 .$$

At resonance the imaginary components cancel out. For the parallel filter the insertion loss is then defined as

$$D = 20 \log [Z/(2 R_p) + 1] . \tag{4}$$

Substituting Eq. (3) into Eq. (4) yields

$$D = 20 \log (1/K+1) . \tag{5}$$

For the series-resonant filter [Figure 1(b)] we find that the equation equivalent to Eq. (2) is

$$1/(\omega_H C_s) - \omega_H L_s = 2 Z + R_s . \tag{6}$$

The resonance condition $\omega^2 L_s C_s = 1$ and the unloaded $Q_u = 1(\omega C_s R_s)$ lead to

$$R_s K = 2Z \, , \tag{7}$$

with K the same as in Eq. (3).

The imaginary components cancel out at resonance and the insertion loss for the series filter is then

$$D = 20 \log[R_s/(2Z) + 1] \, . \tag{8}$$

Substituting Eq. (7) into Eq. (8) leads to a result equal to Eq. (5). Under the original assumption of identical ω, ω_H, and Q_u, K in the series configuration is the same as in the parallel configuration, and hence the insertion loss D is also the same. Since the two circuits are equivalent under the given conditions, only the parallel filter will be considered from now on.

2.2 Tradeoff of Coupling Loss Versus Loaded-Q

For an acceptable bandwidth to be realized, generator and load impedances Z of approximately 50Ω cannot be connected with typical uhf resonant circuits as simply as indicated in Figure 1(a). The L/C ratio of practical devices is such that some form of impedance transformation is required between generator, filter, and load.

Consider Figure 2(a) in which generator and load are connected across a fraction t of the inductance L. The equivalent of this arrangement is presented in Figure 2(b), where generator and load impedances have been transformed by the quantity t^2.

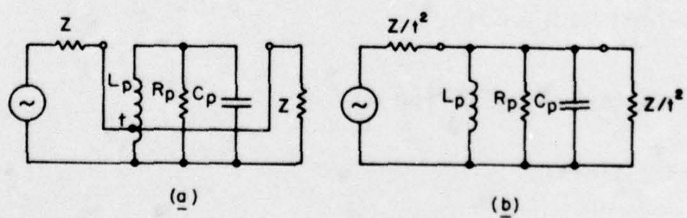

Figure 2. Parallel-resonant Filter With Load Transformation

To investigate the effect of varying degrees of coupling on insertion loss and loaded quality factor we substitute

$$D = 20 \log[Z/(2t^2 R_p) + 1] . \tag{9}$$

for Eq. (4). The unloaded quality factor Q_u for Figure 2(b) is proportional to R_p; the loaded quality factor Q_L is proportional to $1/(2t^2/Z + 1/R_p)$. The ratio of the two is

$$Q_L/Q_u = 1/(2t^2 R_p/Z + 1) . \tag{10}$$

The insertion loss D and the ratio Q_L/Q_u are plotted in Figure 3, with R_p/Z as a parameter. For examples, the D100 curve represents the insertion loss at $R_p/Z = 100$; the Q10 curve represents the Q ratio at $R_p/Z = 10$.

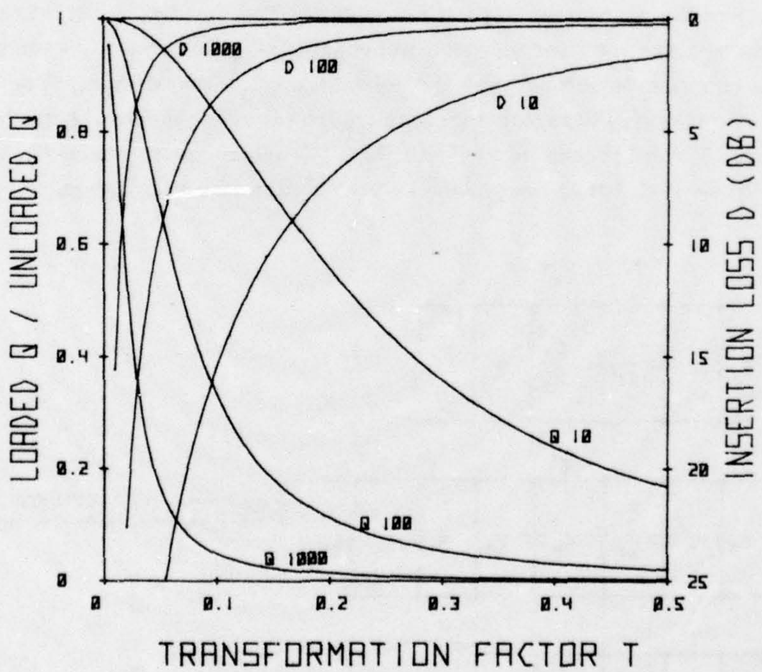

Figure 3. Dependence of Loaded Q and Insertion Loss on Coupling

It is evident from the graph that pairs of curves (D10, Q10; D100, Q100; D1000, Q1000) are similar. As R_p/Z increases, the transformation factor at which a specific combination of Q_L/Q_u and D is reached becomes smaller. Note, for example, the crossover point of the D and Q curves at specific R_p/Z values. Referring now to the D10 and Q10 curves, it is easy to see the tradeoff between insertion loss and quality-factor degradation. Assume 5 dB is the tolerable insertion loss. A reduction in Q to 0.44 of the unloaded value is the consequence. Vice versa, if a Q reduction to 80 percent of the original value is permissible, an insertion loss of 14 dB results. If the receiver output and noise figure are to be affected as little as possible, the unloaded Q of the filter must be very high. The input and output circuit-loading then reduces the filter Q to an acceptable level.

2.3 Varactor-Tuned Parallel-resonant Circuit

For purposes of this investigation, it is sufficient to represent a microwave varactor by a variable capacitor of minimum capacitance C_o, maximum capacitance C_o+C_1, and a series loss resistance R_o. A complete equivalent circuit representation of the loaded helical filter tuned by a varactor is shown in Figure 4(a), where L, R_5, and C are the parameters of the resonator. The quantity $Z/(2t^2)$ is the combined transformed generator and load impedance. Capacitor C_2 is assumed to be a lossless coupling device between the resonator and the varactor. One of the inherent limitations of varactor-tuning at microwave frequencies is the limited varactor Q. A valid technique for improving varactor Q is to connect a lossless capacitor in series, but at the penalty of a reduction in tuning range.

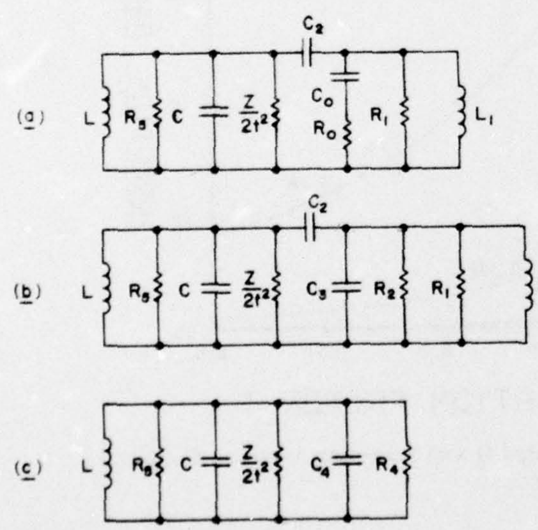

Figure 4. Varactor-tuned Parallel-resonant Circuit

The inductance L_1 is dimensioned to resonate with C_o, the minimum varactor capacitance. Multiplication of L_1 by a factor $B > 1$ allows for partial compensation of C_o and an increase of the resonance frequency of the varactor-loaded resonator toward its unloaded resonance frequency. The losses in L_1 are represented by the parallel resistor R_1.

The initial step in evaluating the resonant network is taken by converting the varactor series elements C_o and R_o to parallel elements. It is necessary to define several other parameters at this point. These are:

- ω resonance frequency of unloaded filter
- ω_o resonance frequency at maximum capacitive load
- ω_1 resonance frequency at minimum capacitive load
- Q quality factor of unloaded filter
- Q_o quality factor at maximum capacitive load
- Q_1 quality factor at minimum capacitive load
- Q_2 quality factor of compensation coil.

From

$$R_o + \frac{1}{j\,\omega_1 C_o} = \frac{1}{\frac{1}{R_2} + j\,\omega_1 C_3} ,$$

it follows that

$$R_2 = \frac{1 + R_o^2\,\omega_1^2\,C_o^2}{R_o\,\omega_1^2\,C_o^2} \tag{11}$$

and

$$C_3 = \frac{C_o}{1 + R_o^2\,\omega_1^2\,C_o^2} . \tag{12}$$

The inductance that resonates with C_3 at ω_1 is

$$L_1 = 1/(\omega_1^2\,C_3) . \tag{13}$$

The losses in L_1 are represented by

$$R_1 = \omega_1\,L_1\,Q_2 . \tag{14}$$

If L_1 is multiplied by a factor B then R_1 is also multiplied by B, assuming that Q_2 is constant.

The parallel resistance of the unloaded resonator is

$$R_5 = Q/(\omega_1 C) . \tag{15}$$

The conversion of the complete loading network consisting of C_2, C_3, R_1, R_2, L_1 in Figure 4(b) into a parallel combination of C_4 and R_4 [Figure 4(c)] follows from

$$\frac{1}{\frac{1}{R_4} + j\omega_1 C_4} = \frac{1}{j\omega_1 C_2} + \frac{1}{j\omega_1 C_3 + \frac{1}{R_2} + \frac{1}{R_1} + \frac{1}{j\omega_1 L_1}} . \tag{16}$$

Equation (16) resolves to

$$C_4 = -\frac{N_o N_2}{\omega_1 (N_1^2 + N_2^2)} \tag{17}$$

and

$$R_4 = \frac{N_1^2 + N_2^2}{N_o N_1} , \tag{18}$$

with

$$N_o = \omega_1 C_2 \left[(\omega_1 C_3 - \frac{1}{\omega_1 L_1})^2 + (\frac{1}{R_2} + \frac{1}{\omega_1 L_1 Q_2})^2 \right] ,$$

$$N_1 = (\frac{1}{R_2} + \frac{1}{\omega_1 L_1 Q_2}) \omega_1 C_2 ,$$

$$N_2 = -(\omega_1 C_2 + \omega_1 C_3 - \frac{1}{\omega_1 L_1})(\omega_1 C_3 - \frac{1}{\omega_1 L_1}) - (\frac{1}{R_2} + \frac{1}{\omega_1 L_1 Q_2})^2 .$$

Note that ω_1, an unknown, can now be determined from the resonance condition

$$\omega_1^2 L(C + C_4) = 1 \tag{19}$$

and Eq. (17).

The loaded Q_1 at ω_1 is found from

$$Q_1 = \omega_1/[2(\omega_1 - \omega_{1H})] ,$$

where ω_{1H} is the halfpower frequency below resonance at which the imaginary and the real admittances are equal. Thus,

$$1/(\omega_{1H}L) - \omega_{1H}(C + C_4) = 1/R_4 + 1/R_5 + 2t^2/Z . \tag{20}$$

Quantities ω_o and Q_o are determined in exactly the same way except that C_o+C_1 are substituted for C_o; L_1 is not recomputed in this case since it was previously selected to resonate with C_o at ω_1 and is now a fixed component during the capacitive tuning. The assumption is made that Q and Q_2 are constant parameters over the tuning range of the filter. Consequently, R_1 and R_5 are proportional to the frequency. The filter is optimized by varying the combinations of t, C_2, and L_1.

Four experimental resonators were built and their parameters ω, Q, and C measured. These are described in Section 3.

3. HELICAL RESONATOR DESIGN

Semiempirical directions for designing coaxial resonators incorporating helical inner conductors have been given by Macalpine and Schildknecht.[2] An idea of the size of the helical resonator for typical values of unloaded Q and resonance frequency can be had from our Figure 5. Macalpine and Schildknecht found that unloaded Q is best when the ratio of the coil to the inside diameter of the shield is about 0.55. Based on this information and the necessity of a compact design, we made the shield of silver-plated brass tubing 2 in. long, with an inside diameter of 1.0625 in. The diameter of the coil, made of 16-ga. was 0.5625 in. To minimize losses in the coil, several wires were wound in parallel for the higher frequencies, when the pitch resulting from fewer turns over the length of the shield was large enough to make this feasible.

The parameters measured on the experimental resonators are listed in Table 1. The resonators, which are numbered 1 through 4 according to their coil structure have identical shields. To determine the unloaded Q (actually the Q obtained with the minimum load necessary for a reliable reading in the measurement setup) the resonator was inserted with low coupling between the generator and load.

2. Macalpine, W.W., and Schildknecht, R.O. (1959) Coaxial resonators with helical inner conductor, Proc. IRE 47(No. 12):2099-2105.

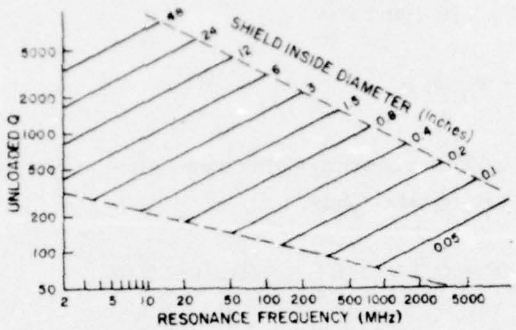

Figure 5. Unloaded Q of Helical Resonators

Table 1. Parameters of Experimental Helical Resonators

Filter Unit No.	1	2	3	4
Number of Turns	12	8	4	2
Parallel Wires	1	3	5	5
$\omega/2\pi$ (MHz)	339	487	919	1420
Unloaded Q	754	1107	892	1206
Equivalent C (pF)	1.03	1.26	1.02	0.58

The equivalent C came from measurements of the change $\Delta\omega$ in resonance frequency when the center of the halfwave resonator was being loaded with capacitance C_p (Figure 6). Note that $C_p/2$ detunes each quarterwave section of the resonator, resulting in

$$C = \frac{C_p}{2\left(\frac{\omega}{\omega - \Delta\omega}\right)^2 - 2} . \tag{21}$$

Capacitance C is the one used in Eq. (15).

The measured unloaded Q is in good agreement with Figure 5 for units 1 and 2. Units 3 and 4 have values lower than predicted but they lie outside the upper frequency boundary suggested in the graph as suitable for helical resonators.

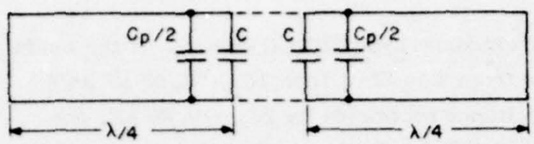

Figure 6. Halfwave Helical Resonator Equivalent Circuit

4. MICROWAVE TUNING-VARACTOR PERFORMANCE

The silicon varactor used in this investigation is typical of high-Q tuning varactors commercially available from various companies. Measured at a reverse bias of 4V, its quality factor exceeds Q = 4000 at 50 MHz. For calculating Q at a particular frequency f, the manufacturer gives the simple relationship $Q = Q_{50MHz} \cdot 50/f\,(MHz)$, which implies that the series loss resistance in the equivalent circuit of the diode is independent of frequency. In a previous report[3] we assumed the series loss resistance to be independent of bias voltage as well, which is justifiable as a first-order approximation. Here we do consider the change in quality factor with bias level. A typical curve of diode Q as a function of bias voltage is shown in Figure 7. The curve is essentially a straight line and fitted to pass through the point $Q_{-4V} = 4000$.

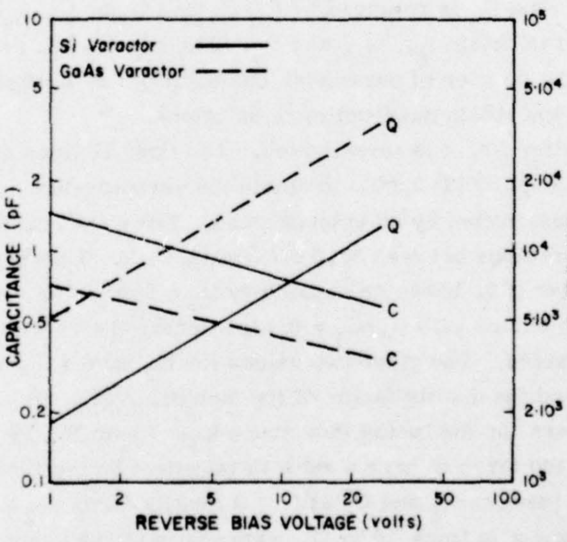

Figure 7. Tuning Varactor Q and Capacitance Versus Bias Voltage

3. Lammers, U. H. W., and Stiglitz, M. R. (1974) Microwave Dielectric Resonator Tuning, AFCRL-TR-74-0497.

The C curve was measured with a Tektronix Type 130 LC-meter. If the useful diode-tuning range is assumed to extend from 1 to 25V, then $2C_o = 0.60$ pF and $Q_3 = 12000$ (at 50 MHz). While the capacitance increases by $2C_1 = 0.86$ pF, the quality factor reduces by $Q_4 = 10200$ (at 50 MHz). Each quarterwave section of the resonator is tuned by C_o or C_o+C_1. Other silicon varactor diodes that were available had higher breakdown voltages and therefore larger ratios of maximum-to-minimum capacitance. This was accompanied by a lower Q and hence not desirable.

5. FILTER OPTIMIZATION BASED ON MEASURED CIRCUIT PARAMETERS

A program (Table 2) was written for the HP9830A/9866A Calculator/Printer to evaluate the resonant coupling network of Figure 4, and to optimize bandwidth and insertion loss for the available resonator and varactor parameters. The program is executed as follows. After defining the input parameters, an iteration is performed to find the resonance frequency f_1 when the resonator is loaded with varactor capacitance C_o. The search starts at the original resonance frequency f of the filter and continues toward the lower frequencies. After establishing f_1, the iterative search is repeated to find the halfpower frequency. The starting point is frequency f_1 this time. With resonance and halfpower frequencies known, Q_1 and D_1 are determined. Next C_o is replaced by C_o+C_1 and Q_3 by Q_3-Q_4. The two iterations are repeated to obtain f_o, Q_o, and D_o. The results are presented in tabular form. Although the number of parameter combinations is limited, it is possible to recognize trends and obtain near-optimum solutions.

In Table 3, filter No. 1 is investigated. The first 12 lines show the results of using a very low coupling (T=0.001) to obtain the varactor-tuned resonator parameters essentially unaffected by an external load. The weak coupling yields insertion losses D_o and D_1 ranging between 37.0 dB and 48.5 dB. This group of 12 cases consists of 4 subgroups of 3, based on capacitors C_2 = 99pF, 1pF, and 0.5pF. A capacitor C_2 = 99pF in series with $C_o+C_1 = 0.73$pF leaves the varactor tuning performance almost unaffected. The other two values for C_2 have a marked effect on both the tuning range and the quality factor of the circuit. Values B = 99, 1, and 2 were tested as multipliers for the tuning inductance L_1. Again B = 99 represents the case where L_1 is too large to have a substantial effect on the filter. For B = 1, resonance occurs between L_1 and C_o at f_1. A quality factor $Q_2 = 200$ for L_1 was used throughout except in lines 10 to 12, where $Q_2 = Q$ was assumed. This is not a realistic assumption since lumped inductors of this quality cannot be realized.

Table 2. Program Determining Filter Tuning Range, Loaded Q, and Insertion Loss

```
10 REM VARACTOR TUNING WITH L COMPENSATION AND SERIES C
20 H=1
30 C=1.03*10↑(-12)
40 C0=0.3*10↑(-12)
50 C1=0.43*10↑(-12)
60 C2=99*10↑(-12)
70 F=0.339*10↑9
80 Q=754
90 Q2=200
100 Q3=12000
110 Q4=10200
120 Z=50
130 T=0.001
140 W=2*PI*F
150 L=1/(W↑2*C)
160 I=0
170 J=0
180 W1=W
190 A=1
200 B=99
210 X0=0
220 N=0
230 IF J=0 THEN 250
240 N=4
250 R0=1/(Q3*PI*10↑8*C0)
260 R2=(1+W1↑2*R0↑2*C0↑2)/(W1↑2*R0*C0↑2)
270 C3=C0/(1+W1↑2*R0↑2*C0↑2)
280 IF J=1 THEN 310
290 IF I=1 THEN 310
300 L1=B/(W1↑2*C3)
310 K=1/R2+1/(Q2*W1*L1)
320 D=W1*C2*((W1*C3-1/(W1*L1))↑2+K↑2)
330 N1=W1*C2*K
340 N2=-K↑2-(W1*C2+W1*C3-1/(W1*L1))*(W1*C3-1/(W1*L1))
350 C4=-N2*D/(W1*(N1↑2+N2↑2))
360 R4=(N1↑2+N2↑2)/(D*N1)
370 R5=Q/(W1*C)
380 G=1/R4+1/R5+2*T↑2/Z
390 X=1-W1↑2*L*(C+C4)
400 IF J=0 THEN 420
410 X=G*W1*(C+C4)-1/(W1*L)
420 N=N+1
430 IF ABSX<10↑(-10) THEN 500
440 IF X-X0=0 THEN 470
450 IF X*X0>0 THEN 470
460 A=A*(-1)
470 W1=W1-A*W/(2↑N)
480 X0=X
490 GOTO 260
500 IF I=1 THEN 610
510 IF J=1 THEN 550
520 F1=W1/(2*PI)
530 J=1
540 GOTO 190
550 Q1=F1/(2*(F1-W1/(2*PI)))
560 D1=20*LGT(Z/2/T↑2*G)
570 I=1
580 C0=C0+C1
590 Q3=Q3-Q4
600 GOTO 170
610 IF J=1 THEN 650
620 F0=W1/(2*PI)
630 J=1
640 GOTO 190
650 Q0=F0/(2*(F0-W1/(2*PI)))
```

Table 2. Program Determining Filter Tuning Range, Loaded Q, and Insertion Loss (Cont)

```
660 D0=20*LGT(Z/2/T↑2+G)
670 K1=10↑12
680 K2=10↑6
690 IF H=0 THEN 750
700 PRINT "  F(MHZ)="F/K2;"Q="Q;"Q3="Q3+Q4;"Q4="Q4
710 PRINT "  C(PF)="C*K1;"C0(PF)="(C0-C1)*K1;"C1(PF)="C1*K1
720 PRINT
730 PRINT "  C2     B      Q2     T       F0     Q0     D0     F1     Q1
740 PRINT "  (PF)                         (MHZ)         (DB)   (MHZ)
750 FORMAT 2F6.1,1F6.0,1F6.3,2F6.0,1F6.1,2F6.0,1F6.1
760 WRITE (15,750)C2*K1,B,Q2,T,F0/K2,Q0,D0,F1/K2,Q1,D1
770 END
```

Table 3. Performance Parameters of Filter 1 (Si varactor)

F(MHZ)= 339 Q= 754 Q3= 12000 Q4= 10200
C(PF)= 1.03 C0(PF)= 0.3 C1(PF)= 0.43

	C2 (PF)	B	Q2	T	F0 (MHZ)	Q0	D0 (DB)	F1 (MHZ)	Q1	D1 (DB)
1	99.0	99.0	200	0.001	260	504	43.1	299	858	37.2
2	1.0	99.0	200	0.001	286	670	39.8	307	849	37.1
3	0.5	99.0	200	0.001	299	741	38.5	312	838	37.0
4	99.0	1.0	200	0.001	295	307	48.5	339	436	44.2
5	1.0	1.0	200	0.001	303	399	45.0	339	435	44.2
6	0.5	1.0	200	0.001	309	484	42.7	339	435	44.2
7	99.0	2.0	200	0.001	276	384	45.9	317	580	41.1
8	1.0	2.0	200	0.001	292	532	42.0	319	615	40.3
9	0.5	2.0	200	0.001	302	629	40.1	321	642	39.8
10	99.0	1.0	754	0.001	295	411	45.9	339	682	40.3
11	1.0	1.0	754	0.001	303	500	43.1	339	682	40.3
12	0.5	1.0	754	0.001	309	574	41.2	339	682	40.3
13	99.0	99.0	200	0.005	260	431	16.5	299	645	11.8
14	99.0	99.0	200	0.010	260	297	7.6	299	364	4.7
15	99.0	99.0	200	0.020	260	133	2.6	299	133	1.4
16	99.0	99.0	200	0.030	260	69	1.3	299	64	0.7

Resonance frequency, loaded quality factor, and insertion loss at maximum or minimum capacitive tuning are listed in columns 5 through 10. When the varactor is directly connected to the resonator (line 1), a 39-MHz tuning range results. The Q_1 = 858 exceeds that of the unloaded resonator because of a high varactor Q_3 at 299 MHz. At the low end of the tuning range, however, the varactor decreases the circuit quality factor to Q_o = 504. Predictably, a series capacitor C_2 = 1pF narrows the tuning range to 21 MHz, and C_2 = 0.5pF leads to a tuning range of 13 MHz. Circuit Q is only somewhat increased at the high end but is substantially increased at the low end of the tuning range in these cases. The Q improvement from line 1 to line 3 is about 50 percent. In lines 4, 5, and 6, C_o is compensated for by L_1, and so the upper end of the tuning range is at the resonance frequency of the unloaded filter. Tuning ranges of 44, 36, and 30 MHz are found together with a substantial reduction in Q.

Two changes were made to lessen the Q-reduction problem when adding a tuning inductance. In lines 7, 8, and 9 the inductance is seen to be increased to twice the value required for resonance at f_1. It appears that L_1, other than shifting the tuning range, does not improve the quality factor—tuning range combination. The 19-MHz range at Q_o = 629 and Q_1 = 642 in line 9 compares favorably within the 13-MHz range at Q_o = 741 and Q_1 = 838 in line 3. Yet the results in line 9 are inferior to those in line 1, the original configuration. It is somewhat amazing to note that using a tuning inductance of the same high Q as the basic filter (lines 10, 11, 12) does not improve the Q_o and Q_1 to approximately their original values. In lines 13 through 16, the parameters are identical with those in line 1 except for an increase in coupling t. The results make it clear that an acceptable filter quality factor cannot be realized without a significant insertion loss.

The other three filters were investigated like the first one. Parameters f, Q, and C for filter No. 2 are listed in Table 4. The direct tuning of the varactor (line 1) yields a range of 50 MHz at Q_1 = 1126 and Q_o = 465. The decreasing varactor quality begins to show in this filter. There is still a small Q_1 improvement at the upper end of the tuning range, but there is definitely a Q_o reduction to less than the original resonator value. On the other hand the Q increase with reduced C_2 is quite obvious. Tuning ranges of 26 and 17 MHz are obtained for C_2 = 1 and 0.5pF, respectively. In the latter case, Q_o closely approaches the original Q. In the rest of the cases the results are qualitatively similar to those obtained with filter No. 1. If some reduction in tuning range is acceptable, line 6 offers an interesting solution: a tuning range of 37 MHz combined with Q values higher than 500. If a lower-loss tuning coil is used, the Q values are improved by up to 100 or more (line 12). A tighter coupling to the resonator in lines 13 to 16 leads to tradeoffs in Q versus D similar to those found in Table 3.

Table 4. Performance Parameters of Filter 2
(Si varactor)

F(MHZ)= 487 Q= 1107 Q3= 12000 Q4= 10200
C(PF)= 1.26 C0(PF)= 0.3 C1(PF)= 0.43

	C2 (PF)	B	Q2	T	F0 (MHZ)	Q0	D0 (DB)	F1 (MHZ)	Q1	D1 (DB)
1	99.0	99.0	200	0.001	388	465	48.3	438	1126	39.6
2	1.0	99.0	200	0.001	422	730	43.7	448	1152	39.2
3	0.5	99.0	200	0.001	438	893	41.6	455	1161	39.0
4	99.0	1.0	200	0.001	431	305	52.9	487	539	46.9
5	1.0	1.0	200	0.001	442	426	49.0	487	539	46.9
6	0.5	1.0	200	0.001	450	548	46.2	487	539	46.9
7	99.0	2.0	200	0.001	408	370	50.7	460	733	43.7
8	1.0	2.0	200	0.001	430	572	46.0	464	794	42.8
9	0.5	2.0	200	0.001	443	734	43.4	466	845	42.1
10	99.0	1.0	1107	0.001	431	401	50.5	487	937	42.1
11	1.0	1.0	1107	0.001	442	531	47.1	487	937	42.1
12	0.5	1.0	1107	0.001	450	653	44.7	487	937	42.1
13	99.0	99.0	200	0.005	388	426	21.1	438	900	13.6
14	99.0	99.0	200	0.010	388	337	11.1	438	553	5.8
15	99.0	99.0	200	0.020	388	184	4.3	438	217	!.8
16	99.0	99.0	200	0.030	388	105	2.2	438	108	0.9

In filter No. 3 (Table 5) the same silicon varactor tunes a resonator whose parameters are f = 919 MHz, Q = 892, and C = 1.02 pF. Direct tuning by the varactor leads to a 106-MHz range but a low-end unloaded Q_o of only 258. For more uniform results across the tuning range, we have to resort to a capacitor coupled in series to achieve (line 3) a range of 36 MHz accompanied by Q values of 588 to 913. None of the cases with B = 1 or 2 produced acceptably high quality factors. The four tests with stronger coupling (lines 13 to 16) show the familiar trend. The different filter units do not show identical insertion losses for a particular t, which is understandable from the differences in filter L/C ratio and Q.

The degradation of varactor performance continues in Table 6, where filter No. 4 (basic parameters f = 1420 MHz, Q = 1206, and C = 0.58pF) is described. The directly coupled varactor produces a tuning range of 207 MHz for an unacceptably low Q_o = 162. In the best case found in this list (line 3), C_2 = 0.5pF reduces the tuning range to 80 MHz and increases Q_o to 430. A further reduction in C_2 appears necessary for acceptable filter performance. From the data presented so far, it is clear that commercially available silicon tuning varactors are a limiting factor in wide-range high-Q tuning performance, particularly at higher uhf frequencies.

Table 5. Performance Parameters of Filter 3
(Si varactor)

F(MHZ)= 919 Q= 892 Q3= 12000 Q4= 10200
C(PF)= 1.02 C0(PF)= 0.3 C1(PF)= 0.43

	C2 (PF)	B	Q2	T	F0 (MHZ)	Q0	D0 (DB)	F1 (MHZ)	Q1	D1 (DB)
1	99.0	99.0	200	0.001	703	258	57.5	809	842	46.0
2	1.0	99.0	200	0.001	773	443	51.9	830	890	45.3
3	0.5	99.0	200	0.001	809	588	49.1	845	913	44.9
4	99.0	1.0	200	0.001	798	184	61.5	919	425	53.0
5	1.0	1.0	200	0.001	820	256	57.5	919	424	53.0
6	0.5	1.0	200	0.001	836	338	54.4	919	424	53.1
7	99.0	2.0	200	0.001	746	216	59.5	858	568	49.9
8	1.0	2.0	200	0.001	791	346	54.3	865	618	48.9
9	0.5	2.0	200	0.001	819	470	51.2	871	661	48.1
10	99.0	1.0	892	0.001	798	219	60.0	919	679	49.0
11	1.0	1.0	892	0.001	820	297	56.2	919	679	49.0
12	0.5	1.0	892	0.001	836	383	53.3	919	679	49.0
13	99.0	99.0	200	0.005	703	250	29.8	809	751	19.0
14	99.0	99.0	200	0.010	703	228	18.6	809	563	9.5
15	99.0	99.0	200	0.020	703	168	9.1	809	281	3.5
16	99.0	99.0	200	0.030	703	118	5.2	809	153	1.7

Table 6. Performance Parameters of Filter 4
(Si varactor)

F(MHZ)= 1420 Q= 1206 Q3= 12000 Q4= 10200
C(PF)= 0.58 C0(PF)= 0.3 C1(PF)= 0.43

	C2 (PF)	B	Q2	T	F0 (MHZ)	Q0	D0 (DB)	F1 (MHZ)	Q1	D1 (DB)
1	99.0	99.0	200	0.001	948	162	61.6	1155	818	45.8
2	1.0	99.0	200	0.001	1081	293	55.3	1202	952	44.1
3	0.5	99.0	200	0.001	1155	430	51.4	1235	1045	43.1
4	99.0	1.0	200	0.001	1164	109	66.8	1420	327	55.6
5	1.0	1.6	200	0.001	1197	144	62.9	1420	326	55.6
6	0.5	1.0	200	0.001	1226	191	59.4	1420	326	55.6
7	99.0	2.0	200	0.001	1038	133	64.1	1266	472	51.4
8	1.0	2.0	200	0.001	1123	213	58.5	1283	536	49.8
9	0.5	2.0	200	0.001	1179	309	54.5	1297	597	48.5
10	99.0	1.0	1206	0.001	1164	129	65.4	1420	610	50.2
11	1.0	1.0	1206	0.001	1198	167	61.5	1420	610	50.2
12	0.5	1.0	1206	0.001	1226	219	58.2	1420	610	50.2
13	99.0	99.0	200	0.005	948	159	33.8	1155	729	18.9
14	99.0	99.0	200	0.010	948	150	22.3	1155	543	9.4
15	99.0	99.0	200	0.020	948	122	12.0	1155	269	3.4
16	99.0	99.0	200	0.030	948	93	7.3	1155	146	1.7

Efforts are under way in industry to produce higher-Q tuning varactors of materials other than silicon. Toward the end of this investigation several developmental units of a new gallium arsenide varactor became available to us. We tested one of these diodes in the four filters in exactly the same way as we did the silicon varactor.

The manufacturer's measured value for the GaAs varactor is Q_{-4V} = 11600. Figure 7 shows a Q curve for the new diode, plotted parallel to the Q curve for the silicon diode and adjusted to pass through Q = 11600 at -4V bias voltage. This is an assumption made for want of any other measured performance data. Based on the extrapolated curve, values of Q_3 = 34000 (at 50 MHz) and Q_4 = 28800 (at 50 MHz) are found. Measured as before, the capacitance - versus - voltage curve gave values of $2C_o$ = 0.34pF and $2C_1$ = 0.36pF. The calculator program was then rerun for all four filters with the new diode parameters.

The GaAs cases 1 (no series capacitor C_2, no tuning inductance L_1, weak coupling) are shown for filters 1 through 4 in Table 7. A tuning range of 21 MHz is achieved in filter 1 at Q_o = 771 and Q_1 = 842. This compares most closely with line 3 of Table 3 (Si), where approximately the same Q values are obtained with a 14-MHz tuning range. Filter 2 when tuned by a GaAs varactor diode covers 26 MHz at Q_o = 926 and Q_1 = 1186. This compares with a 17-MHz tuning range at Q_o = 893 and Q_1 = 1161 (Si, line 3, Table 4). In the case of filter 3 we find a 58-MHz tuning range with Q_o = 617 and Q_1 = 957. The corresponding numbers from line 3, Table 5 (Si) are 36 MHz, Q_o = 588 and Q_1 = 913. Lastly, in filter 4 a tuning range of 127 MHz, Q_o = 466, and Q_1 = 1212 compare with 80 MHz, Q_o = 430, and Q_1 = 1045 in line 3, Table 6 (Si). In all filters there is an approximate improvement of 50 percent in tuning range when comparing the GaAs type with a Si type connected through a 0.5 pF series capacitor.

Table 7. Performance Parameters of Filters 1 to 4 (GaAs varactor)

```
F(MHZ)= 339  Q= 754  Q3= 34000   Q4= 28800
C(PF)= 1.03    C0(PF)= 0.17    C1(PF)= 0.18

C2    B      Q2    T      F0    Q0    D0    F1    Q1    D1
(PF)                      (MHZ)       (DB)  (MHZ)       (DB)
99.0  99.0   200 0.001    293   771 . 38.3   314   842  36.9


F(MHZ)= 487  Q= 1107    Q3= 34000   Q4= 28800
C(PF)= 1.26    C0(PF)= 0.17    C1(PF)= 0.18

C2    B      Q2    T      F0    Q0    D0    F1    Q1    D1
(PF)                      (MHZ)       (DB)  (MHZ)       (DB)
99.0  99.0   200 0.001    431   926   41.4   457  1186  38.8


F(MHZ)= 919  Q= 892  Q3= 34000   Q4= 28800
C(PF)= 1.02    C0(PF)= 0.17    C1(PF)= 0.18

C2    B      Q2    T      F0    Q0    D0    F1    Q1    D1
(PF)                      (MHZ)       (DB)  (MHZ)       (DB)
99.0  99.0   200 0.001    794   617   48.8   852   957  44.4


F(MHZ)= 1420    Q= 1206    Q3= 34000   Q4= 28800
C(PF)= 0.58    C0(PF)= 0.17    C1(PF)= 0.18

C2    B      Q2    T      F0    Q0    D0    F1    Q1    D1
(PF)                      (MHZ)       (DB)  (MHZ)       (DB)
99.0  99.0   200 0.001   1123   466   50.9  1250  1212  41.7
```

6. IMPLEMENTATION OF TUNED HELICAL FILTERS

To validate our calculations we measured the actual tuning performance of helical resonators 1 through 4 when tuned directly by a varactor diode. The experimental setup is shown as a block diagram in Figure 8. A test signal is generated in a doubly balanced mixer by heterodyning a cw signal from the oscillator (0.1 to 2 GHz) with a sweep signal from the POLYSKOP I. The POLYSKOP I (manufactured by Rohde & Schwarz) is a combination test instrument containing a sweep generator (0.5 to 400 MHz in five ranges), a calibrated attenuator, a VTVM, and a TV-tube display. The POLYSKOP I is adjusted to produce a sweep of 40 MHz center frequency and a sweep width of approximately 5 MHz. The lower sideband of the doubly balanced mixer output signal at the R port is tuned with the cw oscillator such that it will cover the passband of the helical resonator under test. For simple and accurate frequency measurements, a frequency-variable marker signal from an

external generator is superposed on the TV-tube trace. The marker-generator and cw oscillator frequencies are measured with the frequency counter. The mixer output signal is coupled into and out of the helical resonator through 3-dB pads. For loose coupling with the resonator, either small coupling loops or capacitive probes are used. When adjusted to the lower sideband the resonator suppresses the upper sideband generated by the balanced mixer. The output from the helical resonator is amplified in a 45-dB or 65-dB wideband amplifier, then rectified by the crystal detector, and displayed on the POLYSKOP I. A dc power supply provides the reverse bias for the varactor through a 1-MΩ decoupling resistor R. The criteria for the selection of R are (1) low voltage crop across R, and (2) negligible reduction in resonator Q.

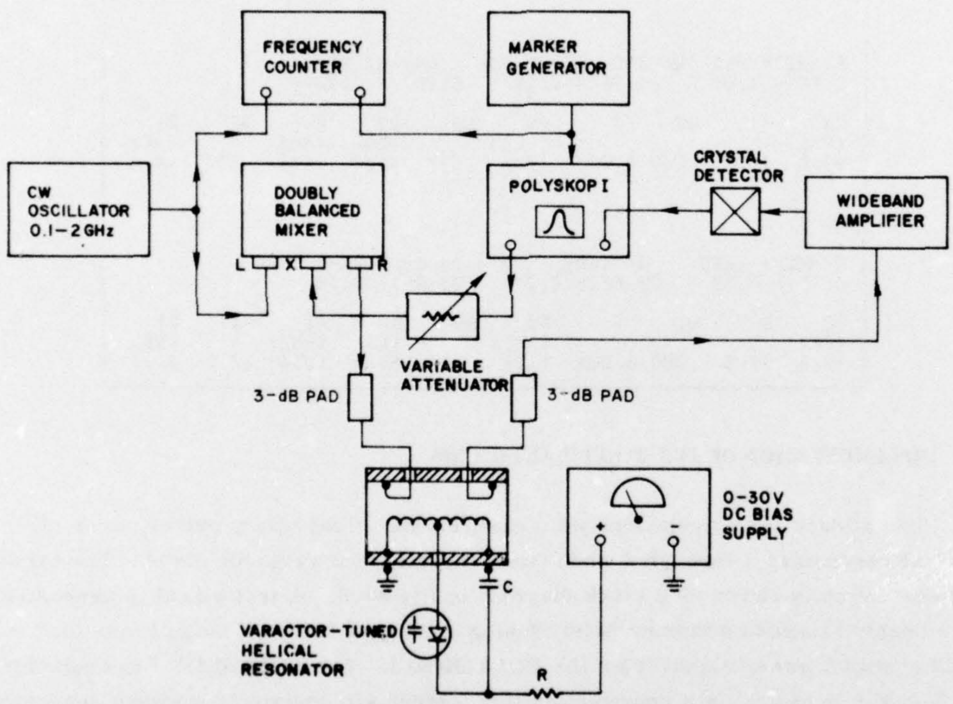

Figure 8. Experimental Setup for Measuring Filter Parameters

Capacitor C = 100pF is large enough not to affect the varactor tuning performance substantially. The physical layout of this capacitor and of a helical resonator is shown in Figure 9. The helix is soldered on both ends to the cylindrical shield. The end plates are removable. A post extends from the center turn of the helix

through the shield. Two-hole flange-mount SMA connectors are fastened directly to the shield. The varactor is mounted externally to the resonator in a small enclosure connected to the capacitor belt strapped around the resonator shield. The external mounting of the varactor is convenient for experiments but is otherwise of no advantage. Insulation between the strap and the shield is provided by 0.005-in. Teflon disks. The bypass capacitor of 100pF is therefore essentially a low-inductance air-insulation type that has high Q. Earlier attempts with a lumped silver-mica capacitor were not successful. Figure 10 is a photograph of a fully assembled unit. Bias voltage is fed in through a coaxial cable. Figure 11 shows all four resonators without the capacitive strap and the end plates. Note the insulated center post.

Calculated and measured results are listed in Table 8 for the silicon varactor and in Table 9 for the gallium arsenide varactor. The calculated data are condensed from Tables 3 through 6 for the Si varactor and from Table 7 for the GaAs varactor.

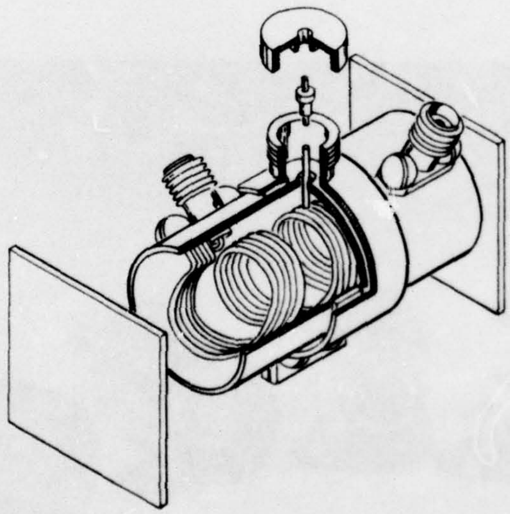

Figure 9. Varactor-tuned Halfwave Helical Filter (cutaway)

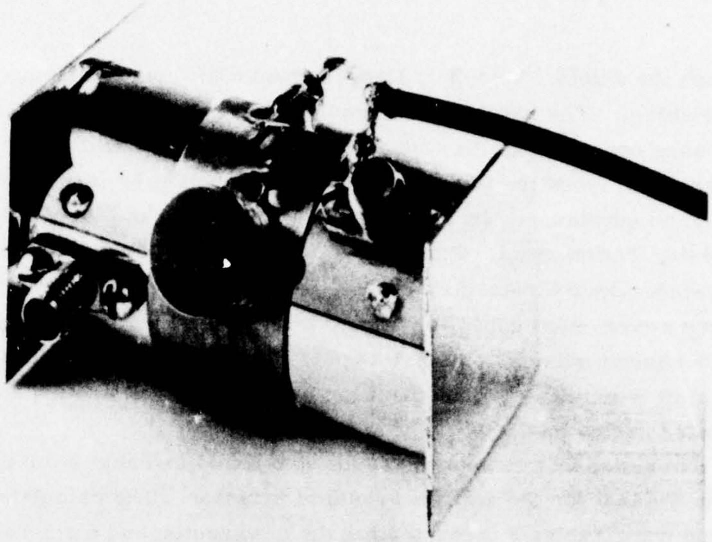

Figure 10. Helical Filter (mounted)

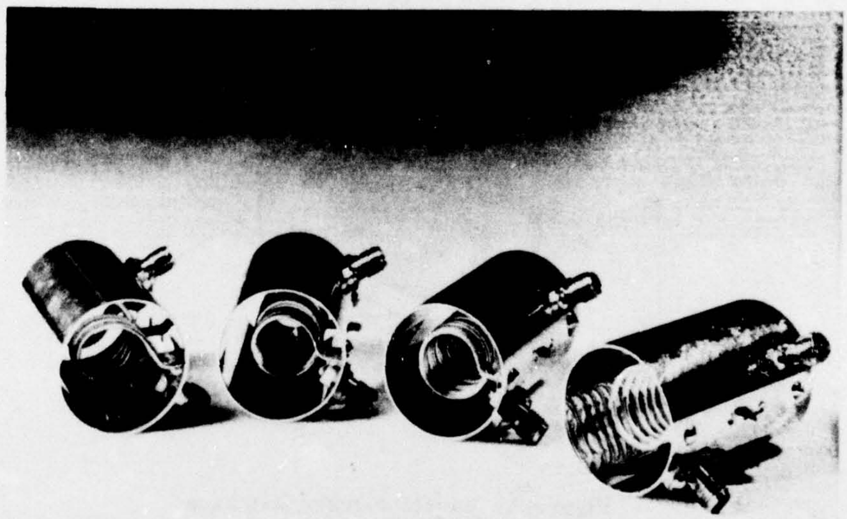

Figure 11. Four Helical Filters Without Tuning Arrangement and End Plates

Table 8. Calculated and Measured Performance of Si Varactor Tuning

Unit	F_o(MHz)	Q_o	D_o(dB)	F_1(MHz)	Q_1	D_1(dB)
1_{calc}	260	504	43.1	299	858	37.2
1_{meas}	262	443	44.0	298	634	49.5
2_{calc}	388	465	48.3	438	1126	39.6
2_{meas}	389	499	41.0	435	855	37.5
3_{calc}	703	258	57.5	809	842	46.0
3_{meas}	701	275	—	807	740	27.0
4_{calc}	948	162	61.6	1155	818	45.8
4_{meas}	949	191	64.0	1164	669	45.5

Table 9. Calculated and Measured Performance of GaAs Varactor Tuning

Unit	F_o(MHz)	Q_o	D_o(dB)	F_1(MHz)	Q_1	D_1(dB)
1_{calc}	293	771	38.3	314	842	36.9
1_{meas}	294	524	48.5	313	728	49.0
2_{calc}	431	926	41.4	457	1186	38.8
3_{meas}	430	729	38.0	454	857	37.5
3_{calc}	794	617	48.8	852	957	44.4
3_{meas}	794	555	26.0	848	771	29.0
4_{calc}	1123	466	50.9	1250	1212	41.7
4_{meas}	1136	460	51.0	1256	937	40.0

The agreement between the calculated and measured frequency range is very good for all filters. The largest discrepancy in frequency is found in unit 4, whose calculated values for the tuning range are 127 MHz and 207 MHz and measured values are 120 MHz and 215 MHz. The error does not appear to be systematic but is attributed to stray capacitances in the experimental setup not accounted for in the calculation. Insertion losses D_o and D_1 vary considerably between theoretical and real values but are not expected to coincide. In the calculations, t = 0.001 was

assumed, which produced a fairly large spread in insertion losses ranging from 61.6 dB to 36.9 dB. On the other hand, mechanically identical coupling loops were used on all filters, which resulted in a range of insertion losses from 64.0 dB to 26.0 dB. The only importance that calculated and measured coupling losses have in this context is to demonstrate that the Q values are for essentially unloaded resonators.

A comparison between calculated and measured Q reveals that the measured quality factor is 11 percent low on the average. This is still considered to be in good agreement with the expected values. It is possible that the external mounting of the diodes is less than optimal. Soldering the diodes directly to the helix might produce slightly better results. The enclosed mounting as sketched in Figure 9 is superior to an open construction previously used, which yielded a much poorer resonator quality factor.

7. CONCLUSIONS

Four tunable halfwave helical resonators covering the frequency range 294 to 1256 MHz were investigated to determine the highest unloaded Q. The varactors used for frequency-tuning are the best low-loss commercial units currently available. Expected and measured filter performance agree well. It was found that at minimum varactor capacitance the varactor Q matches that of the untuned helical structure, that is, with the exception of the Si varactor tuning filters 3 and 4, the varactor is capable of maintaining the overall circuit Q or even improving it somewhat. At maximum varactor capacitance this is true only for the GaAs varactor tuning filter 1. In all other cases the varactor causes a moderate-to-severe Q-degradation of the tuned helical structure, the degradation increasing with frequency. At higher frequencies the varactor is the limiting factor in narrowband low-loss filter design.

At uhf the following parameters were specified for a JTIDS receiver front-end filter:[4]

1) Tuning range: 969 to 1206 MHz
2) Insertion loss: 1 dB (maximum variation: 1 dB)
3) Bandwidth: 10 MHz
4) Group delay distortion: 10 nsec (within 5 MHz from center frequency)

4. Covitt, A.L. (1975) Specifications for Tunable Filter, MITRE Memorandum, 18 November 1975, unpublished.

In Table 10, filter unit 4 with GaAs varactor-tuning was investigated under load conditions, which approximately yield the desired bandwidth. Two parallel varactors (C_o = 0.34pF and C_1 = 0.36pF) were assumed for increased tuning range. Two parallel varactors tuning the halfwave resonators are equivalent to a single varactor tuning a quarterwave resonator. The reason for connecting two varactors in parallel is that larger-capacitance diodes that have very high Q are not available. A tuning range of 171 MHz (16.3 percent) is found in Table 10 which corresponds to 72 percent of the JTIDS specification. Filter 4 was not specifically tailored for this application but serves to illustrate how well the requirements can be met with a single-pole varactor-tuned helical design.

Table 10. Filter 4 Performance With Parallel GaAs Varactors

```
F(MHZ)= 1420    Q= 1206    Q3= 34000    Q4= 28300
C(PF)= 0.58    C0(PF)= 0.34    C1(PF)= 0.36
```

C2 (PF)	B	Q2	T	F0 (MHZ)	Q0	D0 (DB)	F1 (MHZ)	Q1	D1 (DB)
99.0	99.0	200	0.001	959	418	53.3	1130	1263	42.2
99.0	99.0	200	0.040	959	94	2.2	1130	95	0.7

The insertion loss ranges from 0.7 to 2.2 dB and the bandwidth from 10.1 to 11.9 MHz. The group delay distortion within the 3-dB down passband for a single-pole maximally flat time-delay filter becomes 15.8 nsec at 10.1-MHz bandwidth and 13.4 nsec at 11.9-MHz bandwidth. Since this is unacceptably high, a two-pole quarterwave helical filter is considered. With a 3 dB bandwidth of 10 MHz, a maximally flat time-delay filter with two poles will exhibit a group delay distortion of 8.3 nsec. According to Table 10 we assume that an unloaded Q of 1263 can be achieved at 1130 MHz and an unloaded Q of 418 at 959 MHz. The Electronics Engineers' Handbook[5] provides graphs for designing a two-pole filter of finite-Q elements that will produce maximally linear-phase shape with resistive generator and load. Based on these graphs, insertion losses of 0.9 dB at 1130 MHz and 2.8 dB at 959 MHz can be expected for 10-MHz bandwidth.

This puts the two-pole helical filter close to the desired specifications. A low-noise amplifier of ≤ 3-dB gain and wide dynamic range ahead of the filter could be used to compensate for the insertion loss and noise figure degradation.

5. Electronics Engineers' Handbook, Fink, D. G., Ed., McGraw-Hill, 1975.